Alternative Architectures for Narrowband Varactor-Tuned Bandpass Filters

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Abstract— **A unique design approach for narrowband tunable bandpass filters is presented which focuses on engineering the tuned center-frequency dependence of the coupling coefficient between adjacent resonators. New architectures are presented, and equations for the coupling coefficient are derived from energy expressions. A microstrip prototype with independently-tunable center frequency and bandwidth was built and tested, and exhibits excellent performance.**

I. Introduction

The architecture typically used for varactor-tuned bandpass filters is the combline [1], as it is compact, provides excellent stopband performance, and can be designed to have either a constant relative bandwidth or constant absolute bandwidth response. The combline, however, is limited by the fact that a constant absolute bandwidth is difficult to achieve when non-TEM resonators (e.g. microstrip) are used, and that it is not possible to independently tune the center frequency and bandwidth. This paper presents design approaches and alternative architectures aimed at overcoming these limitations.

It is shown in this paper that varactor-tuned transmission-line type resonator structures may be analyzed in an intuitive fashion by assuming a certain distribution of voltage and current, and calculating the coupling coefficient from the energy contained in the resonators and the coupling region. The result is relatively simple expressions for the coupling coefficient versus center frequency of general resonator geometries, which allows for identification of resonator structures with useful tuning properties. In addition, it is shown that the design and optimization of tunable bandpass filters in a circuit simulator is simplified by the identification of coupling resonances between adjacent resonators, and a simple technique for observing these resonances is described.

Finally, it is shown that for certain resonator geometries, using more than one varactor per resonator allows the coupling coefficients (and therefore the bandwidth) to be tuned independently of the center frequency. A microstrip prototype demonstrating this concept was built and tested.

II. Basic Approach

The coupling coefficient between two resonators can be determined from the coupling bandwidth. For the narrowband case:

$$k = \frac{\omega_2^2 - \omega_1^2}{\omega_2^2 + \omega_1^2} \approx \frac{\omega_2 - \omega_1}{\omega_0} \tag{1}$$

$$\omega_0 = \frac{\omega_2 + \omega_1}{2} \tag{2}$$

$$\Delta\omega_{12} = \omega_2 - \omega_1 = \omega_0 k \tag{3}$$

Where k is the coupling coefficient, $\Delta\omega_{12}$ is the coupling bandwidth, ω_1 and ω_2 are the resonant peak frequencies, and ω_0 is the center frequency. The bandwidth of a tunable filter is proportional to $\Delta\omega_{12}$, provided that the shape of the $\Delta\omega_{12}$ vs. ω_0 characteristic is the same for every pair of resonators.

Shown in Fig. 1a is a pair of coupled combline resonators loaded with varactors C_1. This structure can be analyzed using the well-known equivalent circuit of shunt and series shorted stubs [1]. While this approach works well for the combline, the equivalent circuit quickly becomes very complex when the analysis of more general resonator structures is attempted. If the goal is to extract the bandwidth versus center frequency characteristic for more general resonator geometries (and therefore identify other useful tunable filter architectures), a more convenient and intuitive approach is to calculate the coupling coefficient directly from the differential lumped-element coupled-line equivalent circuit (Fig. 1b). This is done by assuming a voltage and current distribution consistent with the boundary conditions in each resonator at resonance, and calculating the energy stored in the resonators and the coupling region. This technique has previously been applied to fixed-tuned coupled resonators [2]. As shown in [2], the capacitive and inductive coupling coefficients k_C and k_L are given by:

$$k_C = \frac{W_{C_m}}{2\sqrt{W_{C_1} W_{C_2}}} \tag{4}$$

$$k_L = \frac{W_{L_m}}{2\sqrt{W_{L_1} W_{L_2}}} \tag{5}$$

where W_{Cm} and W_{Lm} are the total capacitive and inductive energies stored in the coupling region, and W_{C1}, W_{C2} and W_{L1}, W_{L2} are the total capacitive and inductive energies stored in each of the two resonators. These variables are calculated by integrating the capacitive or inductive energy stored in the differential lumped elements. Assuming narrowband conditions, the total coupling coefficient k is:

$$k = \frac{k_L + k_C}{1 + k_L k_C} \approx k_L + k_C \tag{6}$$

and the coupling bandwidth $\Delta\omega_{12}$ is calculated from (3).

Example: Conventional Combline

The approach of calculating coupling coefficients using stored energy is applied to the well-understood combline architecture in order to demonstrate the validity of the

Report Documentation Page

Form Approved
OMB No. 0704-0188

1. REPORT DATE **SEP 2009**	2. REPORT TYPE	3. DATES COVERED **00-00-2009 to 00-00-2009**

4. TITLE AND SUBTITLE **Alternative Architectures for Narrowband Varactor-Tuned Bandpass Filters**	5a. CONTRACT NUMBER
	5b. GRANT NUMBER
	5c. PROGRAM ELEMENT NUMBER
6. AUTHOR(S)	5d. PROJECT NUMBER
	5e. TASK NUMBER
	5f. WORK UNIT NUMBER
7. PERFORMING ORGANIZATION NAME(S) AND ADDRESS(ES) **Naval Research Laboratory,Microwave Technology Branch,Washington,DC,20375**	8. PERFORMING ORGANIZATION REPORT NUMBER
9. SPONSORING/MONITORING AGENCY NAME(S) AND ADDRESS(ES)	10. SPONSOR/MONITOR’S ACRONYM(S)
	11. SPONSOR/MONITOR’S REPORT NUMBER(S)

12. DISTRIBUTION/AVAILABILITY STATEMENT
Approved for public release; distribution unlimited

13. SUPPLEMENTARY NOTES

14. ABSTRACT
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15. SUBJECT TERMS

16. SECURITY CLASSIFICATION OF:			17. LIMITATION OF ABSTRACT	18. NUMBER OF PAGES	19a. NAME OF RESPONSIBLE PERSON
a. REPORT **unclassified**	b. ABSTRACT **unclassified**	c. THIS PAGE **unclassified**	**Same as Report (SAR)**	**4**	

Standard Form 298 (Rev. 8-98)
Prescribed by ANSI Std Z39-18

approach and to provide an example of its use. The distributed-line parameters of the coupled combline resonators shown in Fig. 1b can be written [3]:

$$C' = \frac{1}{2\omega}\left(\frac{y_e}{v_e} + \frac{y_o}{v_o}\right) \quad (7) \qquad L' = \frac{1}{2\omega}\left(\frac{1}{y_e v_e} + \frac{1}{y_o v_o}\right) \quad (8)$$

$$C'_m = \frac{1}{2\omega}\left(\frac{y_o}{v_o} - \frac{y_e}{v_e}\right) \quad (9) \qquad L'_m = \frac{1}{2\omega}\left(\frac{1}{y_e v_e} - \frac{1}{y_o v_o}\right) \quad (10)$$

where y_e and y_o are the even- and odd-mode admittances of the coupled lines, and v_e and v_o are the even- and odd-mode phase velocities. The characteristic admittance of each resonator is:

$$Y = \sqrt{\frac{C'}{L'}} = \sqrt{\frac{y_e y_o (y_e v_o + y_o v_e)}{y_e v_e + y_o v_o}} \qquad (11)$$

Defining:

$$k'_L = \frac{L'_m}{L'} = \frac{y_o v_o - y_e v_e}{y_o v_o + y_e v_e} \quad (12) \qquad k'_C = \frac{C'_m}{C'} = \frac{y_o v_e - y_e v_o}{y_o v_e + y_e v_o} \quad (13)$$

These are the inductive and capacitive coupling coefficients per unit length in the coupling region. Note that $k'_L = k'_C$ when $v_e = v_o$ (TEM). The voltage (V) and current (I) distributions are pure standing waves:

$$V = \sin\theta \quad (14) \qquad I = Y\cos\theta \quad (15)$$

The total inductive energy stored in the resonators and the coupling is then:

$$W_{L_1} = W_{L_2} = \frac{1}{2} L' \int_0^{\theta_0} I^2 d\theta = \frac{1}{2} Y^2 L' \int_0^{\theta_0} (\cos\theta)^2 d\theta \qquad (16)$$

$$W_{L_m} = L'_m \int_0^{\theta_0} I^2 d\theta = Y^2 L'_m \int_0^{\theta_0} (\cos\theta)^2 d\theta \qquad (17)$$

where θ_0 is the electrical length of the resonators as defined in Fig. 1a. From (5), the inductive coupling coefficient is:

$$k_L = \frac{W_{L_m}}{2\sqrt{W_{L_1} W_{L_2}}} = \frac{L'_m}{L'} = k'_L \qquad (18)$$

At resonance the energy stored in the electric field is equal to the energy stored in the magnetic field:

$$W_{C_1} = W_{C_2} = W_{L_1} = W_{L_2} \qquad (19)$$

It is convenient here to use the expressions for inductive energy stored in the resonators when calculating the capacitive coupling coefficient:

$$W_{C_m} = -C'_m \int_0^{\theta_0} V^2 d\theta = -C'_m \int_0^{\theta_0} (\sin\theta)^2 d\theta \qquad (20)$$

$$k_C = \frac{W_{C_m}}{2\sqrt{W_{L_1} W_{L_2}}} = -k'_C \left(\frac{2\theta_0 - \sin 2\theta_0}{2\theta_0 + \sin 2\theta_0}\right) \qquad (21)$$

The total coupling coefficient is sum of capacitive and inductive coupling coefficients according to (6):

$$k = k_C + k_L = \frac{2\theta_0 (k'_C - k'_L) + (k'_C + k'_L)\sin 2\theta_0}{2\theta_0 + \sin 2\theta_0} \qquad (22)$$

In the case of TEM resonators, the coupling coefficient reduces to:

$$k = k_0 \left(\frac{2\sin 2\theta_0}{2\theta_0 + \sin 2\theta_0}\right) \qquad (23)$$

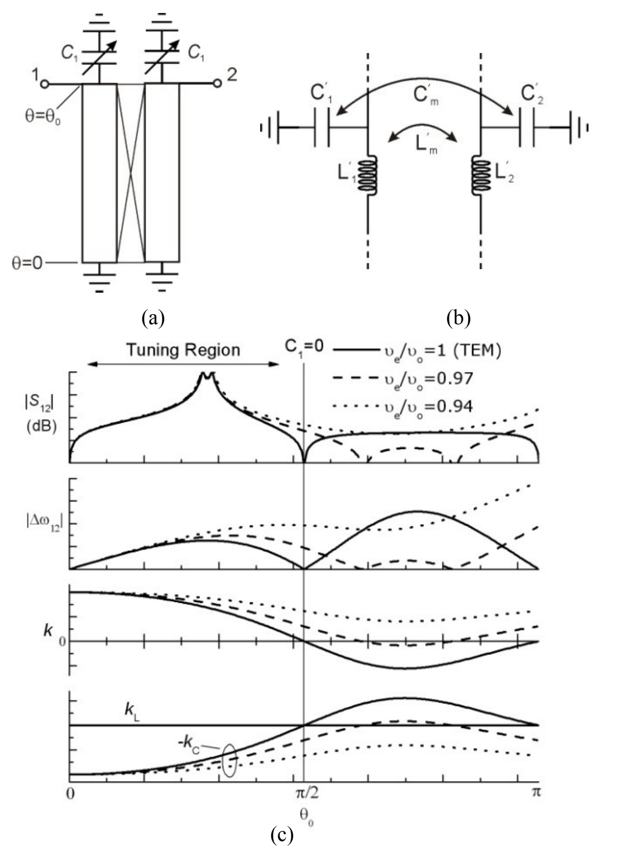

Fig. 1. (a) A pair of coupled combline resonators. (b) Differential coupled-line lumped-element model. (c) Plot of S_{12}, $\Delta\omega_{12}$, k, and k_C, k_L.

where:

$$k_0 = \frac{L'_m}{L'} = \frac{C'_m}{C'} = \frac{y_o - y_e}{y_o + y_e} \qquad (24)$$

Rearranging using trigonometric identities $\Delta\omega_{12}$ can be written:

$$\Delta\omega_{12} = \omega_0 k = \omega_0 k_0 \left(\frac{2\tan\theta_0}{\theta_0 (1 + \tan^2\theta_0) + \tan\theta_0}\right) \qquad (25)$$

which is plotted in Fig. 1c (solid line), with $\omega_0 = \theta_0$. When $\theta_0 = \pi/2$, $C_1 = 0$, which defines the upper limit of the tuning range. The shape of this function is identical to the expression for 3-dB bandwidth given in [1], with a maximum occurring at $\theta_0 = 52.88°$.

III. Coupling Resonances

Shown in Fig. 1c is the simulated S_{12} response (obtained from *AWR* Microwave Office) of a pair of coupled combline resonators (Fig. 1a), the calculated coupling bandwidth $\Delta\omega_{12}$, coupling coefficient k, and inductive/capacitive coupling coefficient (k_C, k_L), all for three different values of v_e/v_o. Note the exact correspondence between the transmission zeroes in the S_{12} response, the zeroes of $\Delta\omega_{12}$, and the zero crossings of k. These frequencies correspond to coupling resonances, where the electric and magnetic couplings between the resonators are equal and out of phase. Proper placement of coupling resonances with respect to the center frequency tuning range allows for an arbitrary bandwidth characteristic

(e.g. constant relative bandwidth or constant absolute bandwidth). The coupling resonances between two resonators can be observed using a circuit simulator by placing two high-impedance ports across the tuning elements in the manner shown in Fig 1a. With the ports placed in parallel with the varactors, the observed coupling resonances remain stationary with varactor tuning. Note that coupling resonances can be present above the realizable tuning range which can be useful for design purposes.

IV. Alternative Architectures

A. Constant Absolute-Bandwidth Architectures

A constant absolute bandwidth is a common system requirement which can be difficult to implement using conventional tunable filter architectures, especially when non-TEM resonators are used. Shown in Fig. 2 are two architectures which are capable of providing a constant absolute bandwidth response with non-TEM resonators. Shown in Fig. 2a is a combline structure where the grounded ends have been decoupled in order to decrease the relative amount of inductive coupling. The expression for the coupling coefficient is calculated using the procedure described in Section II:

$$k = \frac{(k'_C + k'_L)(\sin 2\theta_0 + \sin 2\theta_1) - 2(\theta_0 - \theta_1)(k'_C - k'_L)}{2\theta_0 + \sin 2\theta_0} \quad (26)$$

where θ_1 defines the coupling region. Shown in Fig. 2b is a plot of $\Delta\omega_{12}$ vs. θ_0 for three values of θ_1. The ratio of v_e/v_o is assumed here to be 0.94, a typical value for microstrip. This plot shows that by decreasing the ratio of inductive to capacitive coupling, a constant absolute bandwidth response can be realized when non-TEM resonators are used. This inductive decoupling can be implemented in a number of ways, one of which is with the use of stepped-impedance resonators [4].

Shown in Fig. 2c is a pseudo-combline structure, with the resonators decoupled from the ends where the varactors are connected. The coupling coefficient is given by:

$$k = \frac{2(k'_C - k'_L)\theta_1 + (k'_C + k'_L)\sin 2\theta_1}{2\theta_0 - \sin 2\theta_0} \quad (27)$$

Shown in Fig. 2d is a plot of $\Delta\omega_{12}$ vs. θ_0 for three values of θ_1. Note that when θ_1=0.355θ_0, the $\Delta\omega_{12}$ vs. θ_0 characteristic is extremely flat locally around θ_0=3π/4. Such a structure may be useful for applications where the absolute bandwidth response of conventional combline is insufficiently constant.

B. Architectures which Allow for Independent Control of Center Frequency and Bandwidth

Shown in Fig. 3a is a pseudo-combline structure with varactors C_1 and C_2 placed at opposite ends of the resonators. The resonators are coupled over a region defined by θ_1. Changing the ratio C_1/C_2 effectively varies the ratio of electric to magnetic coupling, and thus the total net coupling and bandwidth. The coupling coefficient of this structure is given by:

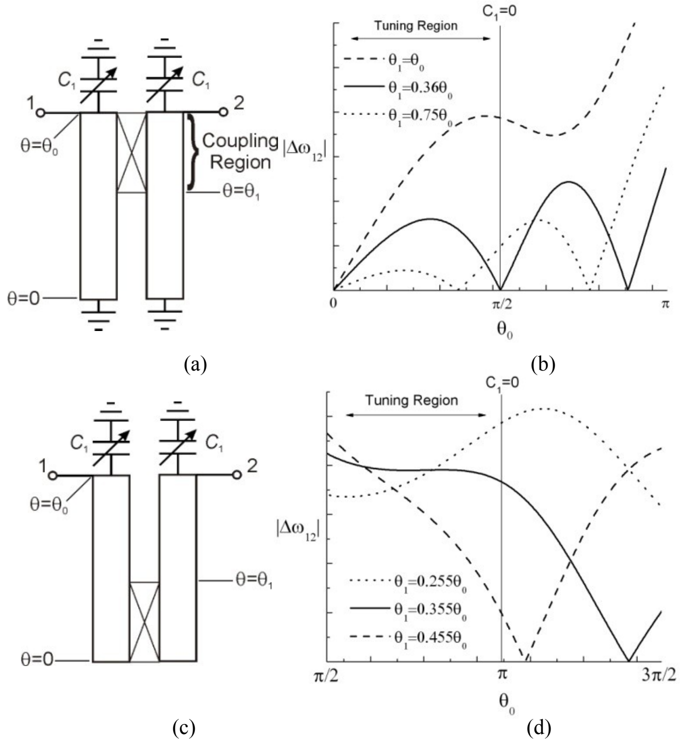

Fig. 2. Constant-absolute bandwidth resonator structures. (a) Combline resonators with reduced inductive coupling. (b) Plot of $\Delta\omega_{12}$ showing effect of decoupling on non-TEM resonators (v_e/v_o=0.94). (c) Pseudo-combline resonators. (d) Plot of $\Delta\omega_{12}$ for three values of θ_1, resonators are assumed to be TEM.

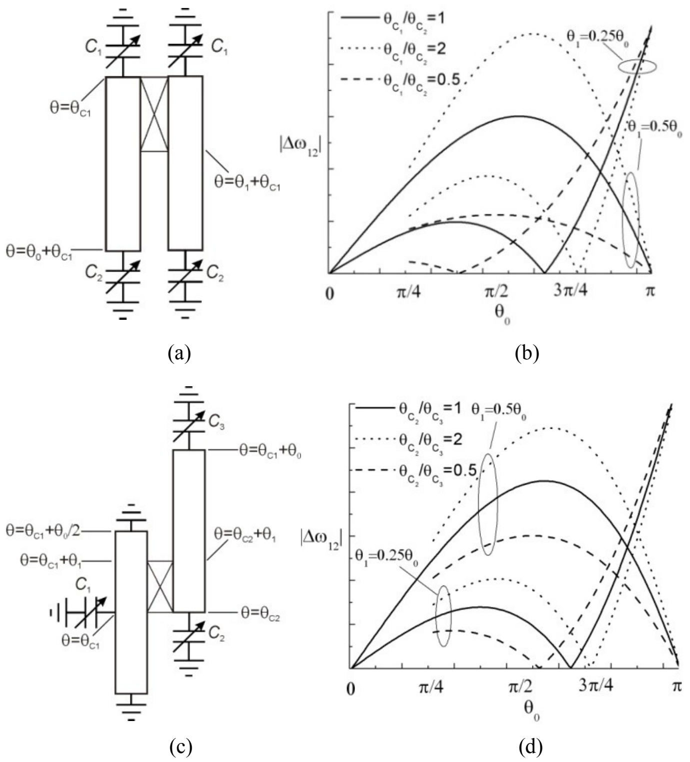

Fig. 3. Resonator structures which allow for independent control of center frequency and bandwidth. (a) Pseudo-combline. (b) Plot showing the effect of differentially tuning C_1 and C_2 on $\Delta\omega_{12}$, for two values of θ_1. (c) Coupled pseudo-combline and grounded pseudo-combline resonators. (d) Plot showing the effect of differentially tuning C_2 and C_3 on $\Delta\omega_{12}$.

$$k = 2\frac{(k'_C - k'_L)\theta_1 + (k'_C + k'_L)\cos(\theta_1 + 2\theta_{C_2})\sin\theta_1}{2\theta_0 + \sin 2\theta_{C_2} - \sin 2(\theta_0 + \theta_{C_2})} \qquad (28)$$

Plotted in Fig. 3b are two sets of $\Delta\omega_{12}$ vs. θ_0 curves, for $\theta_1 = 0.25\theta_0$ and $\theta_1 = 0.55\theta_0$. Each set contains three curves corresponding to different values of θ_{C1}/θ_{C2}, where θ_{C1} and θ_{C2} are the equivalent electrical lengths of C_1 and C_2 ($C_{1,2}$=tan $\theta_{C1,2}$).

Shown in Fig. 3c is a mixed resonator structure consisting of a pseudo-combline and grounded pseudo-combline resonator. The coupling coefficient for this structure is:

$$k = \frac{(k'_C + k'_L)\sin\theta_1 \sin(\theta_0/2 - \theta_1 - \theta_{C_2}) + (k'_C - k'_L)\sin(\theta_0/2 + \theta_{C_2})\theta_1}{\sqrt{(\theta_0 + \sin\theta_0)(2\theta_0 + \sin 2\theta_{C_2} - \sin 2(\theta_0 + \theta_{C_2}))}} \qquad (29)$$

Plotted in Fig. 3d are the $\Delta\omega_{12}$ vs. θ_0 characteristics for various values of θ_1 and θ_{C2}/θ_{C3}, demonstrating the bandwidth-tuning capability of this structure.

It has been shown that by differentially tuning two varactors connected to the same resonator, the ratio of electric and magnetic coupling can be controlled, and therefore the total coupling and bandwidth. This approach to independently tuning bandwidth and center frequency has the advantage of not degrading the stopband unlike other methods such as using intermediate tuning elements [5].

V. RESULTS

A 3rd-order microstrip prototype was built using the resonator geometry shown in Fig. 3c. This particular architecture was chosen as it provides for a symmetric passband response (the coupling between non-adjacent resonators is minimized) as well as a narrow form factor. The substrate is *Rogers* RO4003 (ε_r=3.38, thickness=60 mils) and the varactors are *Aeroflex Metelics* MGV-125-24 (GaAs, hyper-abrupt, C_j = 0.35-7.3 pF). To maintain a good return loss response (~20 dB) with bandwidth tuning, the input and output couplings are tuned using varactor-loaded open-ended stubs. Shown in Fig. 4 is the fabricated circuit as well as plots of the measured center frequency tuning and bandwidth tuning. At a bandwidth of 40 MHz, the center frequency tuning range extends well over an octave. The bandwidth at the center of the tuning range is tunable from 120 MHz down to 0 MHz (the filter is effectively switched off).

VI. CONCLUSIONS

A tunable bandpass filter design approach focused on engineering the frequency dependence of the coupling coefficient between resonators has been presented. The method of determining the coupling coefficient from the calculation of stored energy has been successfully applied to tunable filters. It is observed that an efficient design approach is to identify coupling resonances in a circuit simulator. A microstrip prototype was designed, built, and measured and gives excellent performance.

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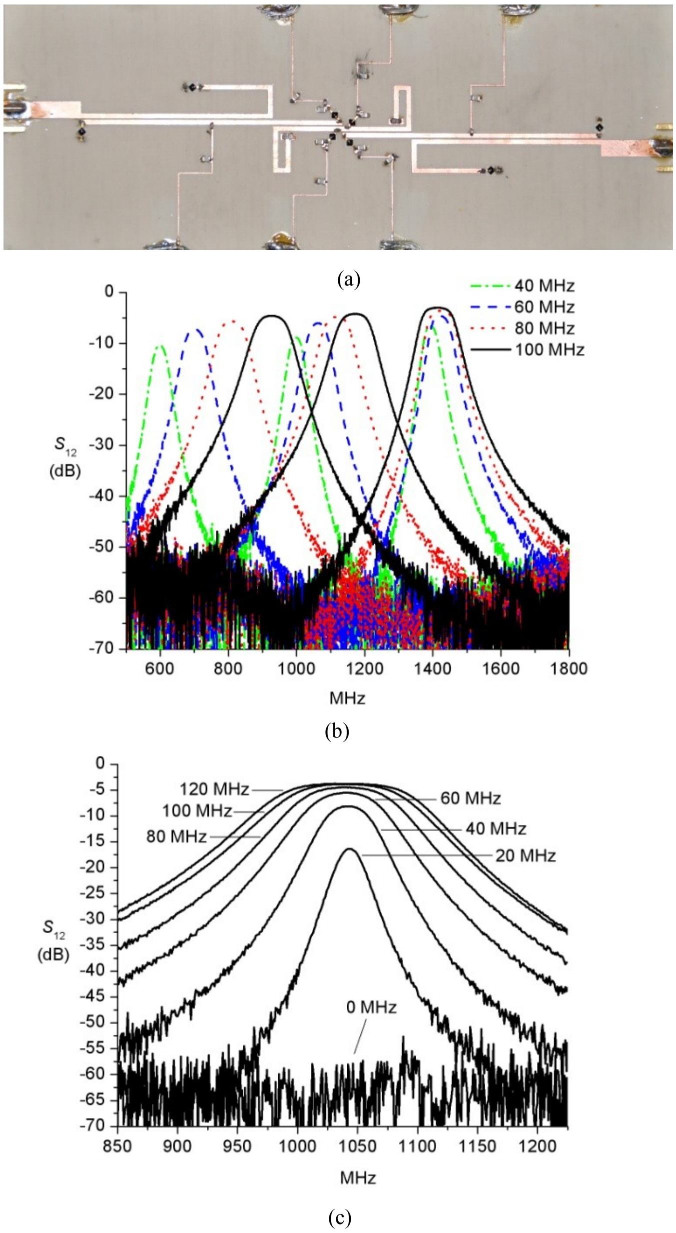

Fig. 4. 3rd-order microstrip prototype with independently tunable center frequency and bandwidth. (a) Fabricated circuit. The central resonator is a pseudo-combline with grounded ends, with two varactors in the center. Outer resonators are pseudo-combline with varactors at both ends. Input and output return loss is tuned with the use of varactors placed at the ends of the input and output transmission lines. (b) Plot of measurements showing center-frequency tuning range for various bandwidths. (c) Plot of measurements showing bandwidth tuning only.